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(71) Applicant(s)
British Broadcasting Corporation

(Incorporated in the United Kingdom)

Broadcasting House, LONDON, W1A 1AA, United Kingdom

(72) Inventor(s)
Mark Charles Dobell Maddocks
Jonathan Highton Stott

(74) Agent and/or Address for Service
Reddie & Grose
16 Theobalds Road, LONDON, WC1X 8PL, United Kingdom

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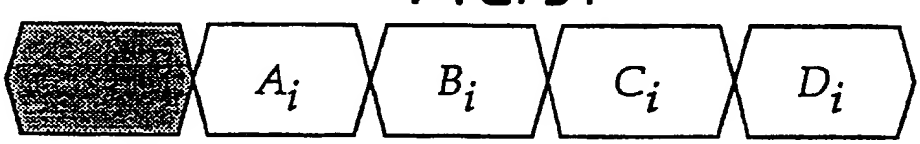
(54) **Coding an orthogonal frequency division multiplexed (OFDM) signal**

(57) An orthogonal frequency division multiplexed (OFDM) signal is coded by dividing each OFDM symbol into a plurality of mini-symbols. Carriers, A, B, C, D, in each mini-symbol are formed from different portions of a plurality of signal sets J, K, L, M. The most important data is coded on a signal set which provides a substantially constant contribution to the corresponding carriers in each mini-symbol and is thus conveyed most ruggedly. Less important data is coded on signal sets which vary between corresponding carriers from mini-symbol to mini-symbol in dependence on predetermined functions and is thus conveyed less ruggedly. The contributions from these signal sets sum substantially to zero across an OFDM symbol.

Alternatively the signal sets coded with the most important data are assigned every rth carrier forming an OFDM symbol and signal sets coded with less important data are assigned to intermediate carriers wherein each intermediate carrier is coded with data derived from a plurality of signal sets.

In each of these systems the most important data serves as a phase and amplitude reference for the signal sets carrying less important data.

FIG. 5.



$A_i =$	$B_i =$	$C_i =$	$D_i =$
$+J_i$	$+J_i$	$+J_i$	$+J_i$
$+K_i$	$+K_i$	$-K_i$	$-K_i$
$+L_i$	$-L_i$	$-L_i$	$+L_i$
$+M_i$	$-M_i$	$+M_i$	$-M_i$

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At least one drawing originally filed was informal and the print reproduced here is taken from a later filed formal copy.

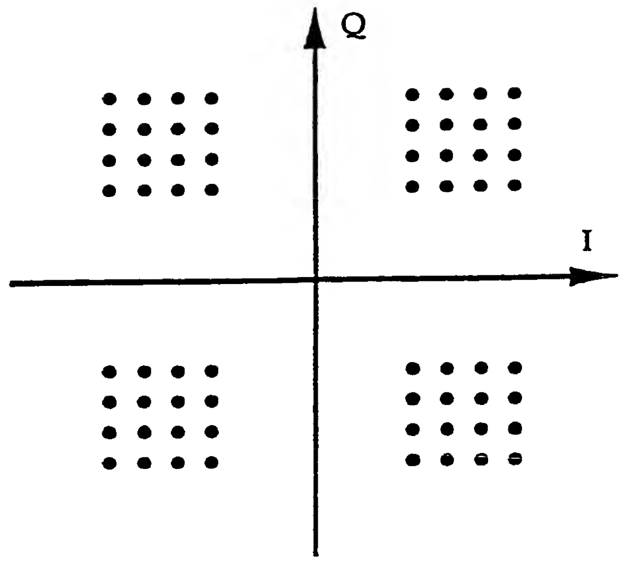


FIG. 1.

FIG. 2.

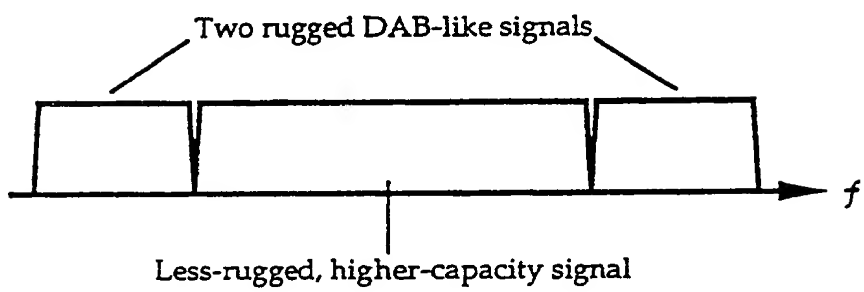
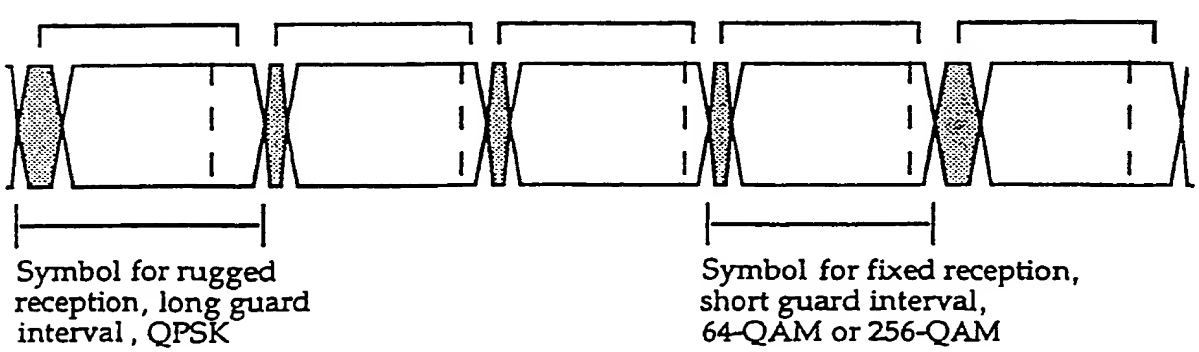


FIG. 3.



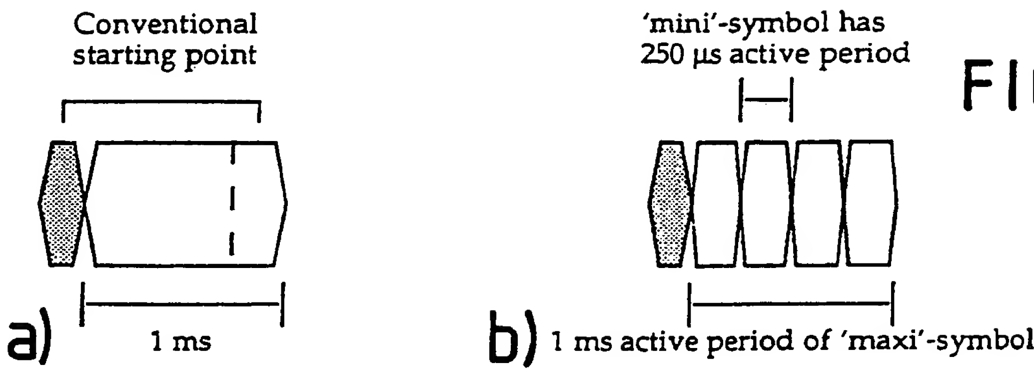


FIG. 4.

FIG. 5.

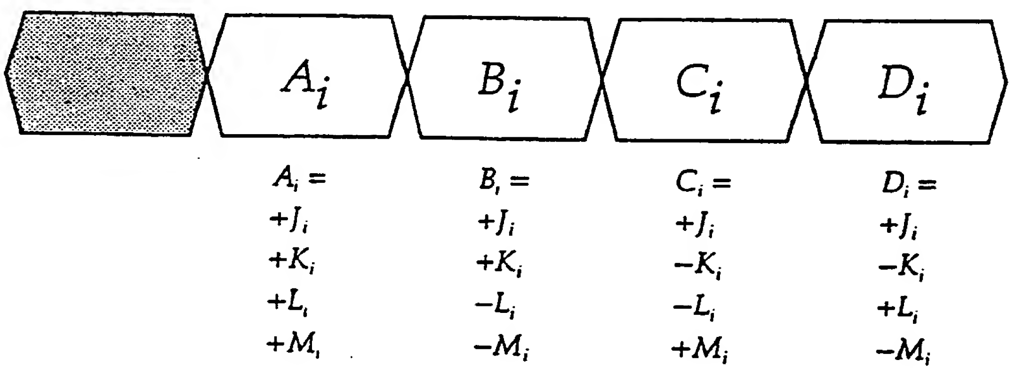
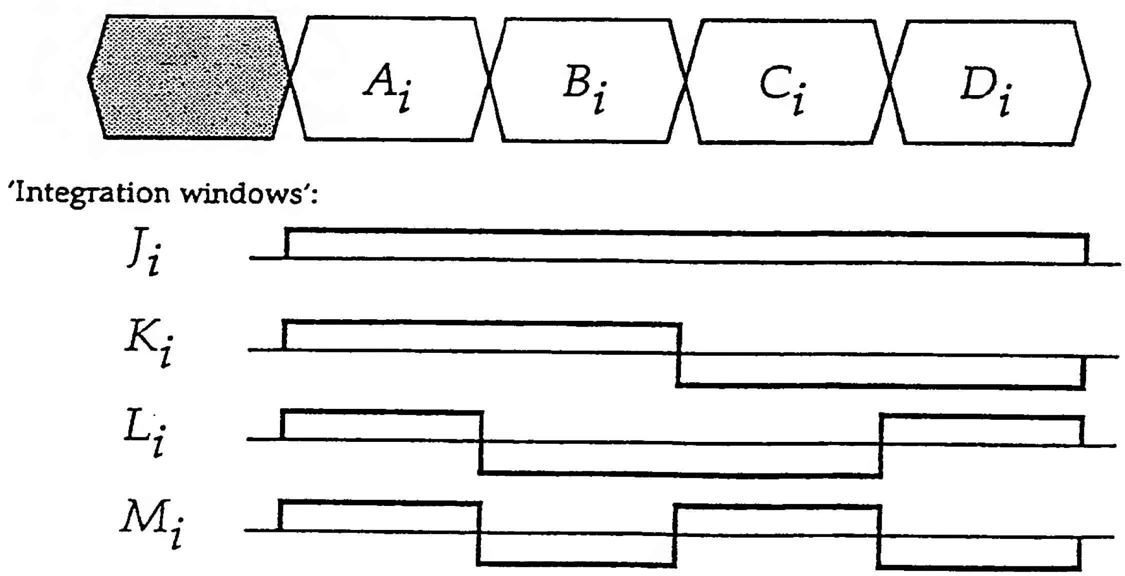


FIG. 6.



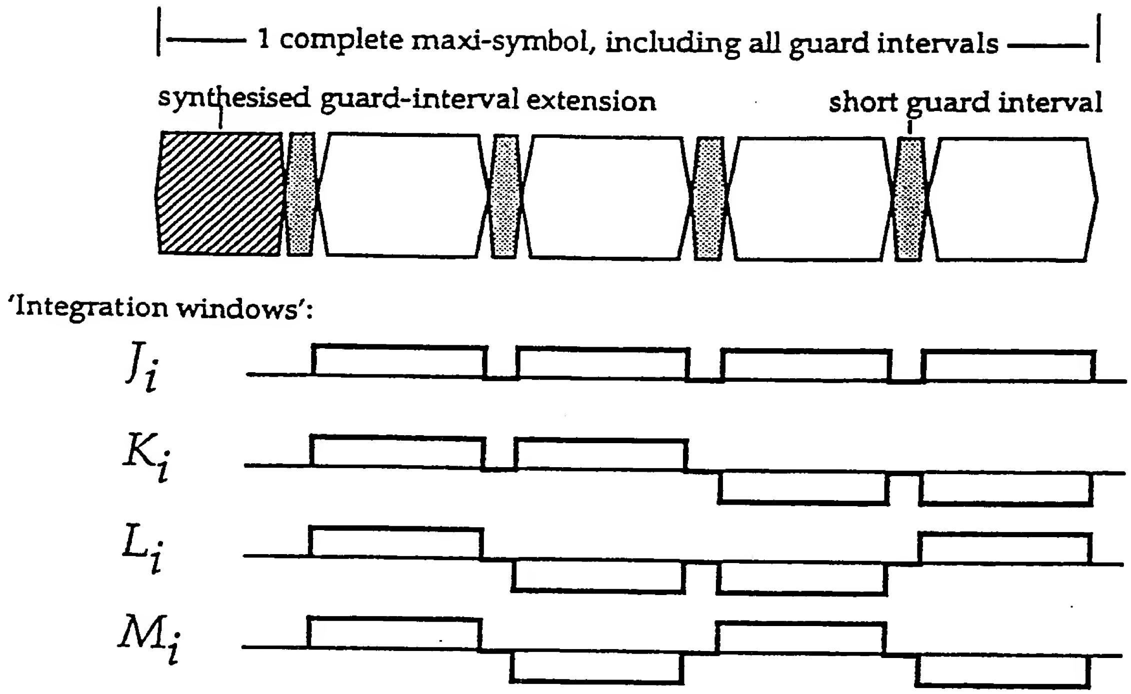

1 complete maxi-symbol, including all guard intervals
synthesised guard-interval extension
short guard interval
'Integration windows':
J_i
K_i
L_i
M_i

FIG. 7.

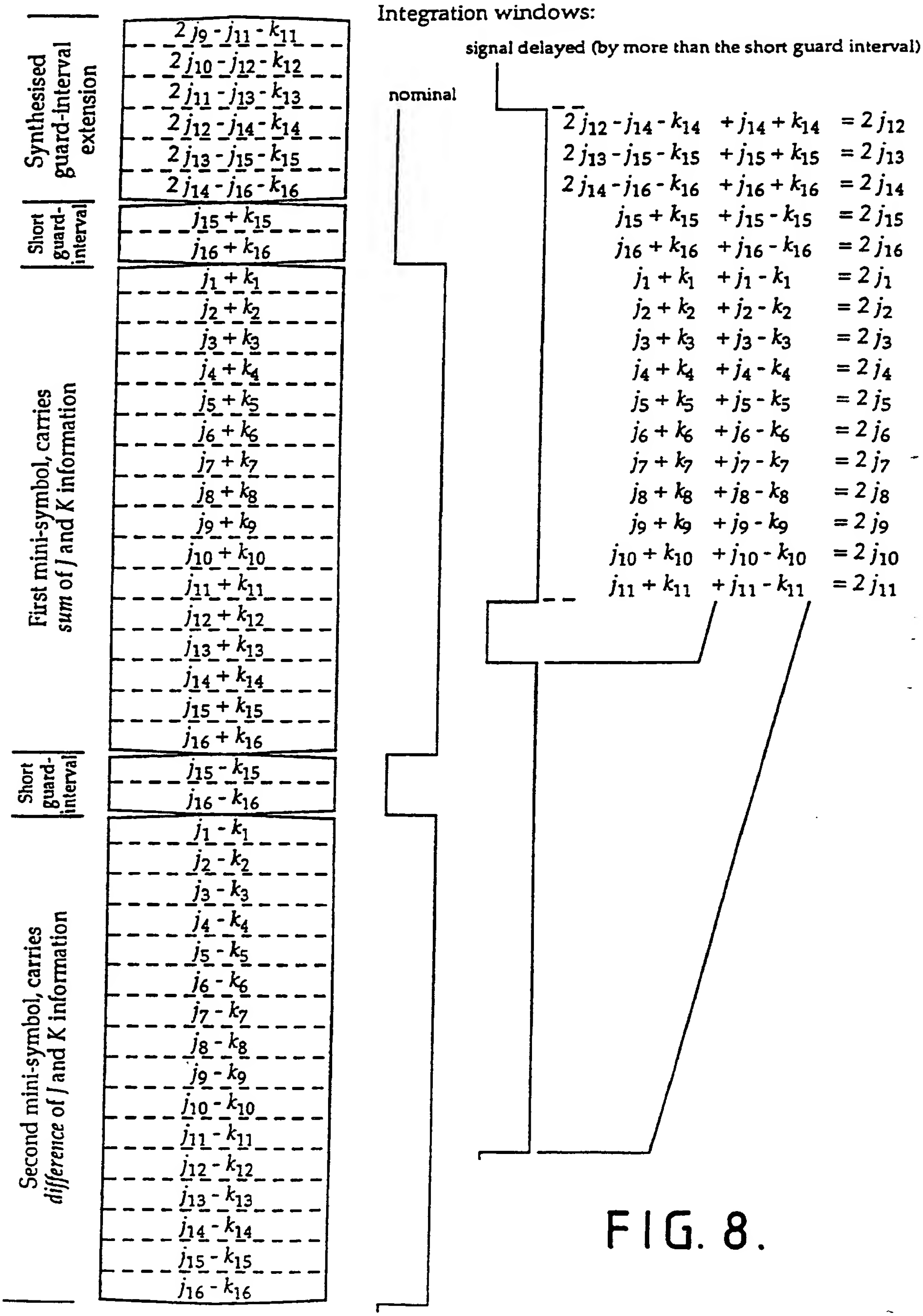

Integration windows:
signal delayed (by more than the short guard interval)
nominal
Synthesised guard-interval extension
Short guard-interval
First mini-symbol, carries *sum* of *J* and *K* information
Short guard-interval
Second mini-symbol, carries *difference* of *J* and *K* information
$2j_9 - j_{11} - k_{11}$
$2j_{10} - j_{12} - k_{12}$
$2j_{11} - j_{13} - k_{13}$
$2j_{12} - j_{14} - k_{14}$
$2j_{13} - j_{15} - k_{15}$
$2j_{14} - j_{16} - k_{16}$
$j_{15} + k_{15}$
$j_{16} + k_{16}$
$j_1 + k_1$ … $j_{16} + k_{16}$
$j_{15} - k_{15}$
$j_{16} - k_{16}$
$j_1 - k_1$ … $j_{16} - k_{16}$
$2j_{12} - j_{14} - k_{14} \quad + j_{14} + k_{14} \quad = 2j_{12}$
$2j_{13} - j_{15} - k_{15} \quad + j_{15} + k_{15} \quad = 2j_{13}$
$2j_{14} - j_{16} - k_{16} \quad + j_{16} + k_{16} \quad = 2j_{14}$
$j_{15} + k_{15} \quad + j_{15} - k_{15} \quad = 2j_{15}$
$j_{16} + k_{16} \quad + j_{16} - k_{16} \quad = 2j_{16}$
$j_1 + k_1 \quad + j_1 - k_1 \quad = 2j_1$
$j_2 + k_2 \quad + j_2 - k_2 \quad = 2j_2$
$j_3 + k_3 \quad + j_3 - k_3 \quad = 2j_3$
$j_4 + k_4 \quad + j_4 - k_4 \quad = 2j_4$
$j_5 + k_5 \quad + j_5 - k_5 \quad = 2j_5$
$j_6 + k_6 \quad + j_6 - k_6 \quad = 2j_6$
$j_7 + k_7 \quad + j_7 - k_7 \quad = 2j_7$
$j_8 + k_8 \quad + j_8 - k_8 \quad = 2j_8$
$j_9 + k_9 \quad + j_9 - k_9 \quad = 2j_9$
$j_{10} + k_{10} \quad + j_{10} - k_{10} \quad = 2j_{10}$
$j_{11} + k_{11} \quad + j_{11} - k_{11} \quad = 2j_{11}$

FIG. 8.

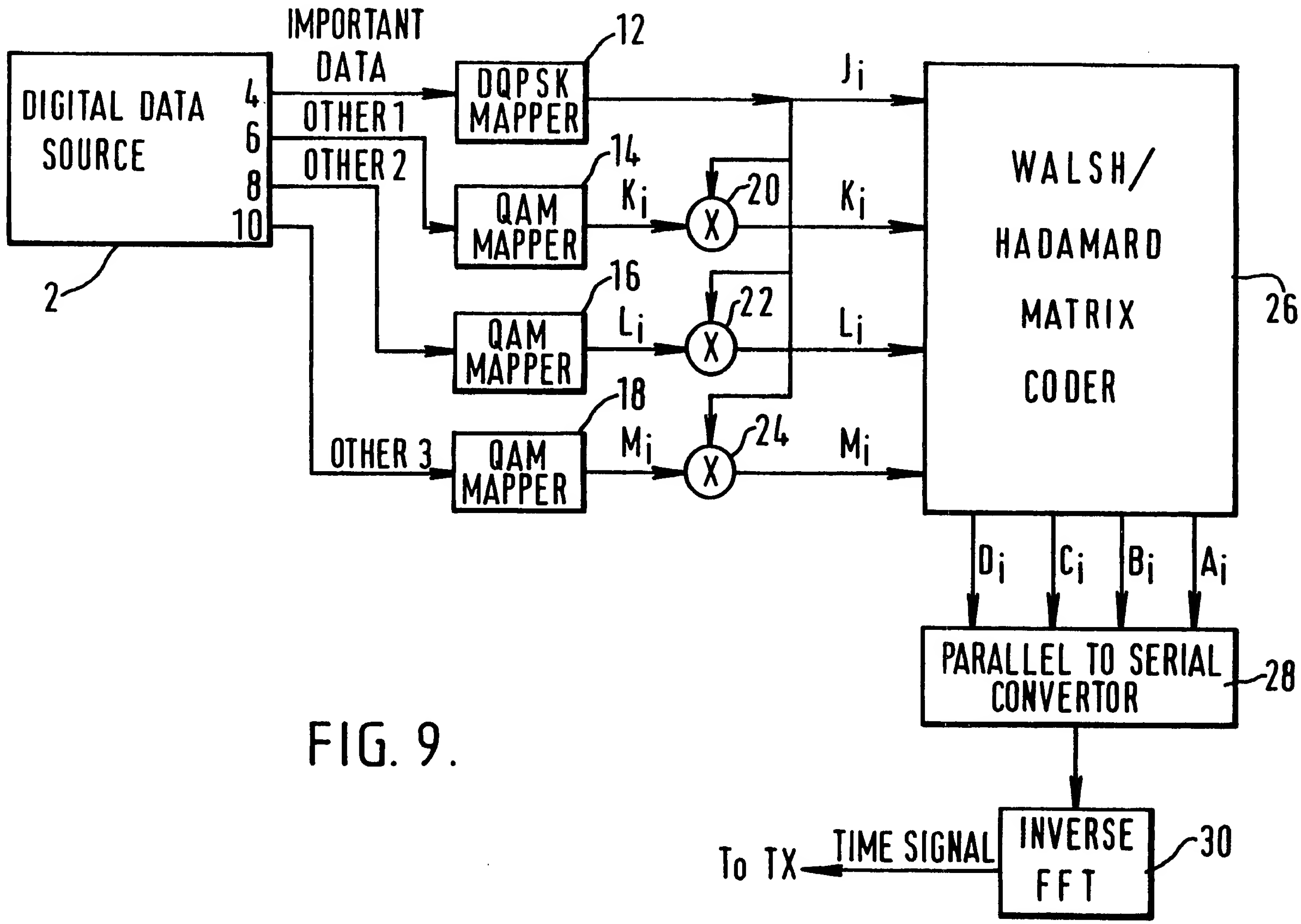
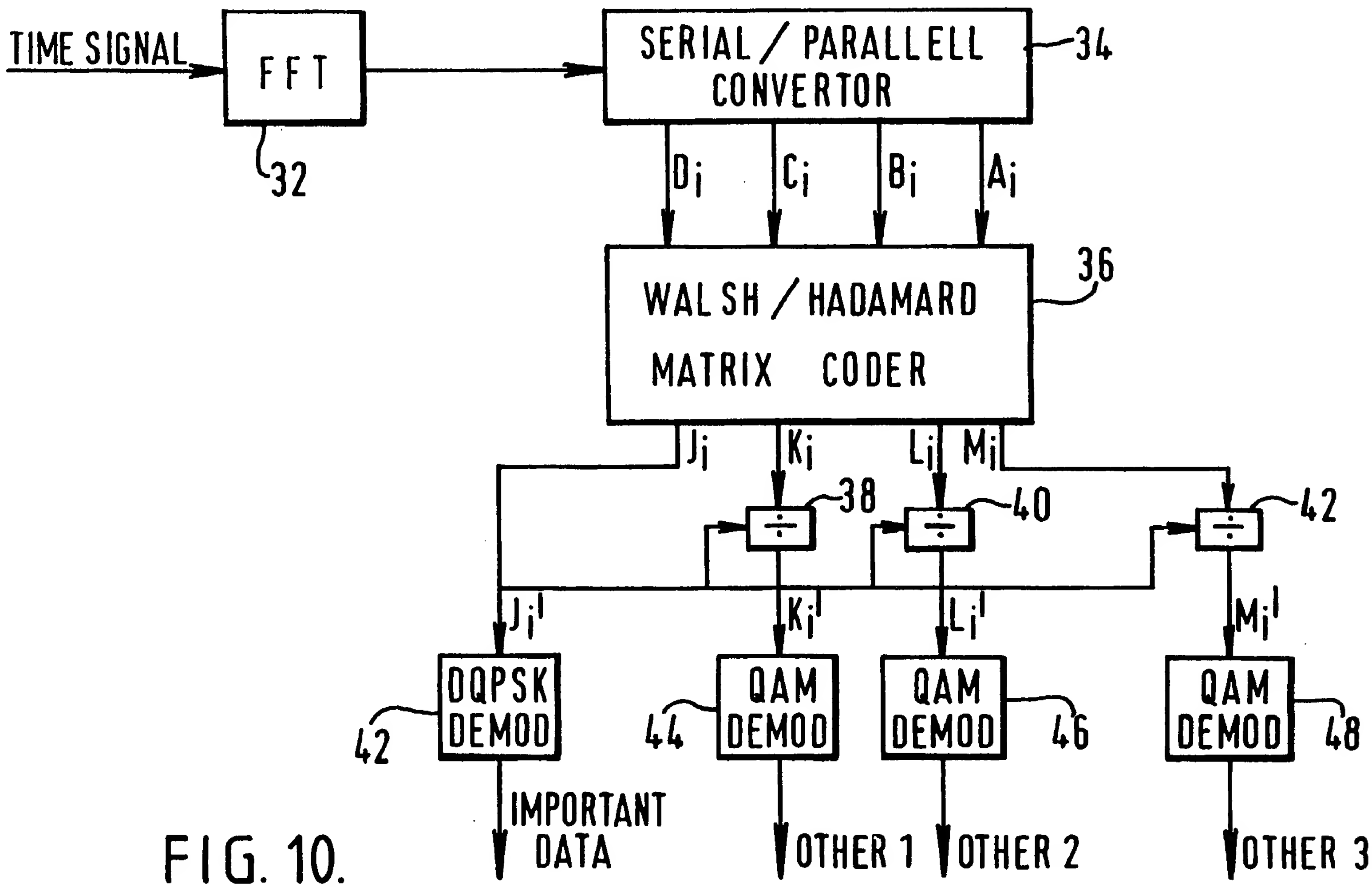

IMPORTANT DATA
DIGITAL DATA SOURCE
2
4
6
8
10
OTHER 1
OTHER 2
OTHER 3
DQPSK MAPPER
12
QAM MAPPER
14
QAM MAPPER
16
QAM MAPPER
18
K_i
L_i
M_i
X
20
X
22
X
24
J_i
K_i
L_i
M_i
WALSH/ HADAMARD MATRIX CODER
26
D_i
C_i
B_i
A_i
PARALLEL TO SERIAL CONVERTOR
28
INVERSE FFT
30
TIME SIGNAL
To TX

FIG. 9.



TIME SIGNAL
FFT
32
SERIAL / PARALLELL CONVERTOR
34
D_i
C_i
B_i
A_i
WALSH / HADAMARD MATRIX CODER
36
J_i
K_i
L_i
M_i
38
40
42
J_i^I
K_i^I
L_i^I
M_i^I
DQPSK DEMOD
42
QAM DEMOD
44
QAM DEMOD
46
QAM DEMOD
48
IMPORTANT DATA
OTHER 1
OTHER 2
OTHER 3

FIG. 10.

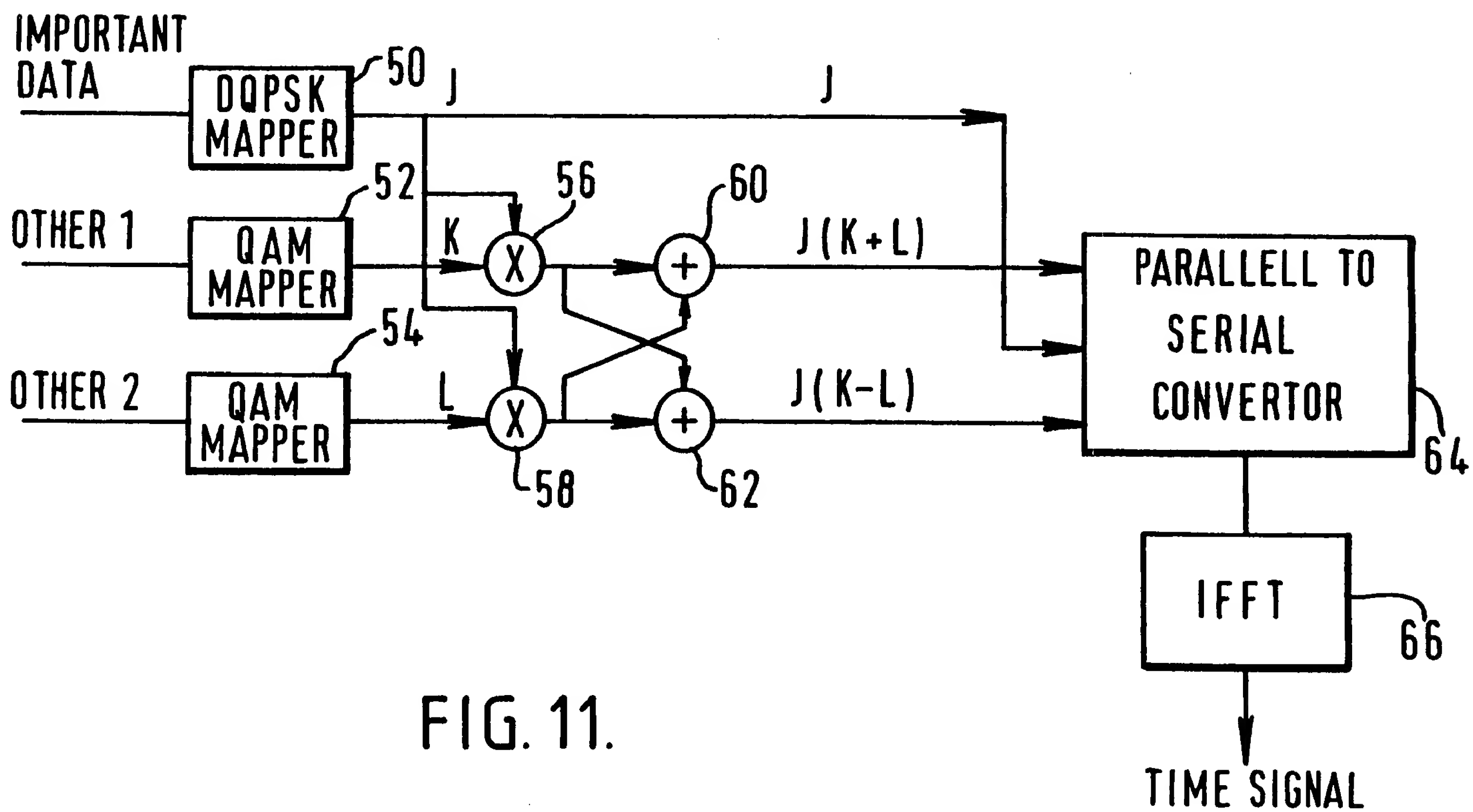

IMPORTANT DATA
DQPSK MAPPER
50
J
J
OTHER 1
QAM MAPPER
52
K
56
60
J(K+L)
OTHER 2
QAM MAPPER
54
L
58
62
J(K-L)
PARALLELL TO SERIAL CONVERTOR
64
IFFT
66
TIME SIGNAL

FIG. 11.

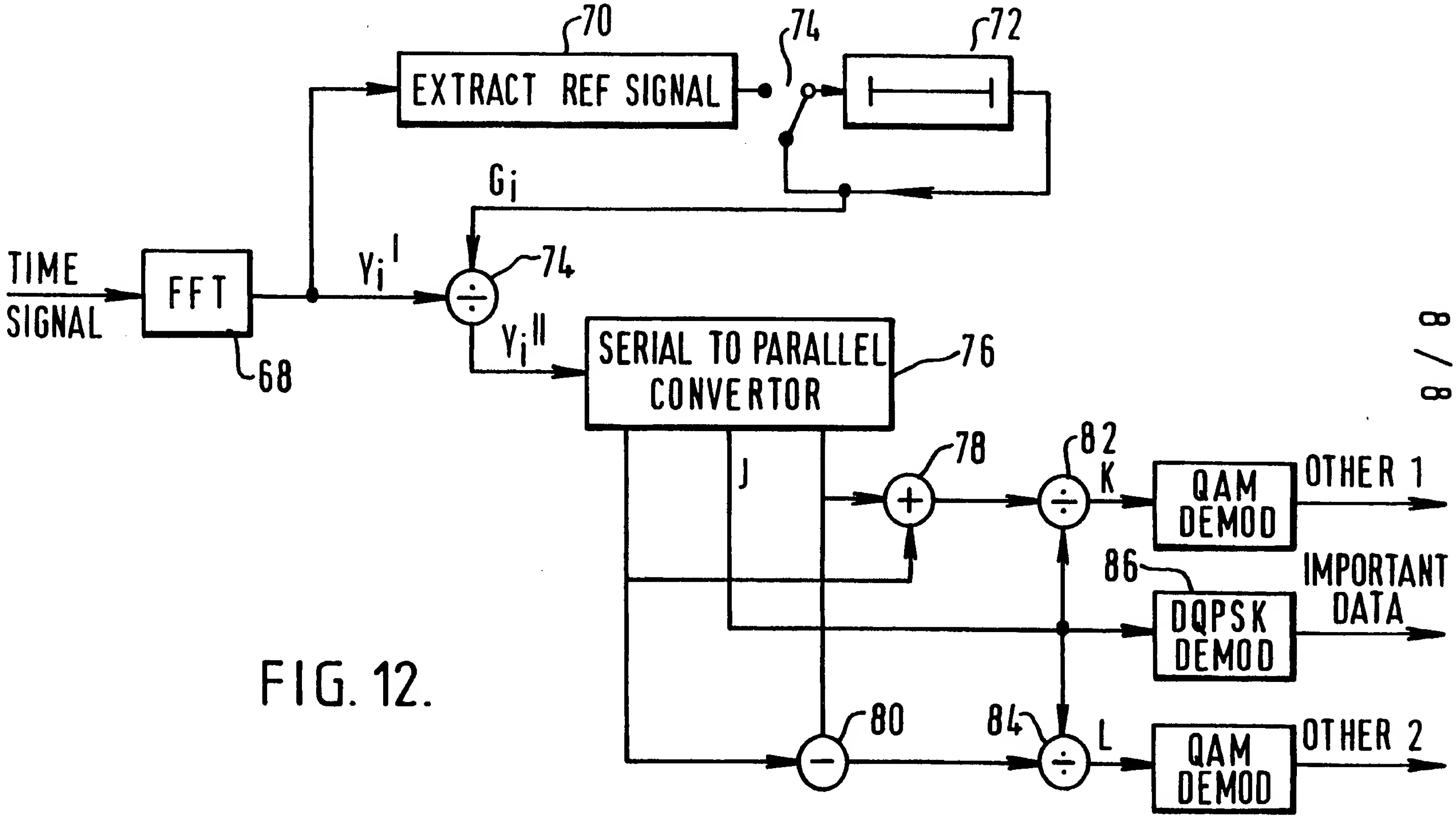

TIME SIGNAL
FFT
68
EXTRACT REF SIGNAL
70
74
72
G_i
$Y_i{}^{I}$
74
$Y_i{}^{II}$
SERIAL TO PARALLEL CONVERTOR
76
J
78
80
82
K
84
L
86
QAM DEMOD
DQPSK DEMOD
QAM DEMOD
OTHER 1
IMPORTANT DATA
OTHER 2

FIG. 12.

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IMPROVEMENTS TO DIGITAL TRANSMISSION SYSTEMS

This invention relates to digital transmission systems and in particular to orthogonal frequency division multiplexed (OFDM) transmission systems which are suitable for terrestrial digital video broadcasts (T-DVB).

In our British Patent Application No. 9313987.1 there is proposed a system in which two or more OFDM signals coded by quadrature amplitude modulation (QAM) are overlaid. The most important data is coded on a large amplitude (e.g quadrature phase shift keying (QPSK) QAM with a low bit rate. To this is added a lower amplitude QAM constellation (e.g.16 QAM) carrying, at a higher bit rate, and therefore at much lower ruggedness, data of lesser importance. A phase diagram illustrating this is shown in Figure 1. It can be seen that the four points of the original QPSK constellation have been turned into "clouds" around which the points of the 16 QAM constellation are arranged. Clearly the amplitude for higher bit rate signals must be kept fairly small so that degradation of the low bit rate data is minimised. The purpose of the system is to enable a simple receiver e.g. a portable receiver to decode the basic signal from the important QPSK coded data whilst better quality receivers with fixed antennas are able to improve the basic signal with the additional 16 QAM coded data.

We have appreciated that there are other ways in which an OFDM signal may be split into portions of data of more and lesser importance carried by more and less rugged portions of a transmission. Various proposals for this are described below.

A first proposal relates to frequency divisional multiplexing (FDM) of different modulation systems. In this the transmitted OFDM ensemble is arranged such that certain carriers are more rugged than others. The ruggedly modulated carriers can convey all of the timing and synchronising information needed to demodulate the less rugged data. Also they can act as channel sounding carriers to provide information needed to remove the effects of amplitude and phase variations on

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the less rugged carriers.

An extension of this idea is permissable if the ensemble is split to comprise two or more disjoint sets of carriers (with-small guard frequencies between the disjoint sets). The idea is to use different guard intervals on the rugged and less rugged carriers. This allows the rugged carriers to operate in a single frequency network (SFN).

The second proposal is time divisional multiplexing (TDM). In this the OFDM symbols are divided in time such that some are more rugged than others. By doing this the guard interval for the rugged and less rugged symbols can be chosen independently.

The third proposal is to use a doubly orthogonal approach. This may be implemented in time or frequency division modulation. This has the advantage that phase and amplitude references needed to demodulate the less ruggedly conveyed information can be extracted from the more ruggedly conveyed information carried in the same time step. This should yield considerable simplification in receiver circuitry as well as improving immunity to phase noise and channel variation.

These various proposals are described below with reference to the accompanying drawings in which:

Figure 1 is the prior art system referred to above:

Figure 2 shows a frequency division multiplexed signal;

Figure 3 shows a time division multiplexed signal;

Figure 4a) and b) show symbols for a doubly-orthogonal coded signal embodying the invention;

Figure 5 shows the complex carrier components for the mini-symbols of Figure 4;

Figure 6 shows the integration windows for each of the components sets of Figure 5;

Figure 7 shows a modification of the system of Figures 4, 5 and 6 including guard intervals;

Figure 8 shows schematically the synthesis of

the guard interval extensions of Figure 7.

Figure 9 shows an embodiment of an encoder for use in a transmitter for the doubly orthogonal system;

Figure 10 shows a decoder for a receiver compatible with the transmitter of Figure 9;

Figure 11 shows an alternative embodiment of a doubly orthogonal coder; and

Figure 12 shows a decoder compatible with the encoder of Figure 11.

FREQUENCY DIVISION

We propose sending two signals with DAB (digital audio broadcast) like characteristics at the extremes of the available channel with another signal sandwiched in between as shown in Figure 2. The DAB signals are of course COFDM based on QPSK with differential demodulation. The active symbol period is 1ms, giving OFDM carriers with 1 kHz spacing. The guard interval is 256μs and would thus permit single-frequency network operation. Each DAB signal has 1536 carriers, and thus occupies a bandwidth of 1.536 MHz. Allowing for error coding each carries - 1.43 Mbit/s of data, which in this case is intended for rugged reception by portables.

An obvious advantage is that portables can use technology already developed for DAB, giving high confidence that it will work. The size of the FFT and the sample rate at the FFT input are the same as for DAB (digital audio broadcast), although in this case all symbols in a frame would have to be processed rather than just one sound-programmes's worth as in DAB. Nonetheless the proposal should reduce development costs for receiver manufacturers.

The 'meat' in the sandwich is the signal intended to be received in addition by 'quality' receivers equipped with rooftop antennas. It would use a higher-order modulation system such as 64-QAM or 256-QAM. It could in principle be either single-carrier (SC) or OFDM. In the SC case an equaliser would probably be necessary.

If the 'meat' uses OFDM then it would be

sensible to synchronise the symbols transmitted in it and the DAB slices surrounding it. When they are co-timed and of the same length (and thus have the same carrier spacing) and the carriers all lie on the same regular raster there is no interference between the 'meat' and the 'DAB slices' and no guard band is necessary. AT the transmitter all the components could be formed up in one operation using a large FFT although the receiver (especially the portable) can pick them off separately using a smaller FFT. (That for each 'DAB' part is the same, known-to-be-practicable size as for DAB; that for the 'meat' might be of double that size or greater).

Now we come to a simple observation which may nevertheless be profoundly useful.

The 'DAB' receiver for each of the 'slices' is able to measure the channel impulse response (and via DFT the complex frequency response) in the manner well-known for DAB. It can do this unambiguously even when echoes occur with greater delay than the guard interval as long as their delay is somewhat less than half the active symbol period. Since the same physical objects (gasholders, remote transmitters in an SFN) cause the echoes seen by both 'DAB' receivers they should determine the same result for the magnitude of the impulse response. The complex values will differ because of the different phases with which they may be received.

When the greatest delay is no more than the guard interval this measurement (which effectively samples the frequency response at every carrier in the 'DAB' signals) is in fact over-sampled. This suggests that these measurements made in two blocks at the edges of the channel might be enough to determine the frequency response over the 'meat' in the middle. Unfortunately there is probably insufficient information to specify completely the correction which must be applied to the 'meat' signal. For example a very short-delay echo which caused a notch in the middle of the 'meat' while scarcely affecting the DAB blocks would probably not be resolved accurately enough. If the 'meat' uses OFDM then the

methods of channel estimation already known are probably simplest and should be applied to measure it directly.

But if the 'meat' is a SC signal then the 'DAB' measurements of impulse response would provide an accurate starting point for an adaptive equaliser. The need to transmit a training sequence might be avoided. An FIR equaliser for a SC system which was designed to deal with a single predominant long-delay echo has been proposed. It used no training sequence but performed an exhaustive search (under microcomputer control) to locate the echo and then adapted the equaliser to deal with it. It worked well but the search did take time which we could avoid with the foreknowledge given by the measurements made on the 'DAB' signals.

Remember that a 'fixed' receiver making use of this 'meat' signal has the benefit of a directional antenna, so hopefully it won't suffer much from neutral multipath. Nor should it suffer much from the unnatural multipath of an SFN. The obvious exception is where another, more distant transmitter lies 'behind' the wanted transmitter on the same bearing. A similar exception occurs with natural multipath and a reflecting object behind the transmitter. Both of these will give rise to the sort of single-predominant-echo impulses response which an FIR equaliser should be able to deal with. The echo amplitude must be rather less than the main signal as an FIR equaliser gives imperfect correction. An IIR equaliser in principle can correct any echo amplitude but at the price of the risk of instability and growth of quantising noise.

TIME DIVISION

Instead of dividing up the channel into frequency blocks using different modulation schemes we could divide it up in time. Some symbols would use a rugged modulation scheme for all carriers simultaneously while others used a less rugged one. The symbols could be of different length.

If we assume that in SFN operation the 'plugs-

free' receiver, with its omni-directions antennas, must be able to cope with long-delay echoes of substantial strength then the signal it receives must have a long guard interval. This implies symbols which are longer still if the guard interval is not to represent a great waste of capacity.

Symbols which are only intended to be received by 'rooftop' antennas could perhaps make do with shorter guard intervals, if we accept the argument that antenna directivity reduces the amplitude of echoes substantially. (There remains the potential problem of the worst scenario of two transmitters on the same bearing). If we shorten the active symbol length of these symbols in the same proportion as the guard interval then the capacity is unchanged from a system where all symbols are the same length. Nothing is gained, except some symbols can be demodulated with a smaller FFT - not very useful.

If however we shorten only the guard interval of the less-rugged symbols then there is a modest gain in capacity as shown in Figure 3.

In DAB the guard interval is one-quarter of the active symbol period and so the capacity is reduced by the factor 0.8. If we keep this guard-interval length for the rugged symbols, but a much shorter one for the others then the factor will be increased, depending on the proportion of less-rugged symbols. We could gain about 10%.

DOUBLY-ORTHOGONAL CODING

This idea combines the orthogonal basic functions (sine waves) of ODFM with the use of orthogonal Walsh-Hadamard type functions to combine data which is transmitted in groups of adjacent mini-symbols. The capacity is the same as for time-division, frequency-division or combined time/frequency-division. Its advantage is that the FFT size is reduced and it also appears to offer a possible solution to the problem of deriving a phase and amplitude reference for demodulating the high-order modulation of the less-rugged part of the signal intended for 'rooftop' reception.

Suppose we form an active symbol of only one-quarter the length, e.g. 250μs as shown in Figure 4. If we keep the same signal bandwidth (and sampling rate) we now have one-quarter of the carriers (n/4, say), and one-quarter of the FFT size. Now suppose we transmit four of these mini-symbols in succession, to occupy the same duration as our original active symbol length. Call this a mini-symbol.

Let each mini-symbol convey the same information, with the same amplitude and phase, using a rugged modulation, say QPSK. Each carrier will be continuous at the join between symbols. In fact there is no difference between this signal and one which was generated as a single symbol of mini-symbol length in which only one carrier in four is sent (with the same amplitudes and phases as those of our mini-symbols) while the others are set to zero.

We can receive this signal by correlating it with each carrier frequency; for each we integrate-and dump over the whole maxi-symbol period. This could be done by using a full size FFT over the whole period - it would yield n/4 finite answers and 3n/4 zero results for the inbetween carriers. Since 3n/4 results will be zero anyway we could use a decimated transform. Imagine instead we perform the smaller FFT for each mini-symbol and sum the results, carrier by carrier. This will give an identical result. In the absence of noise each mini-symbol will give the same answer anyway; when noise is added we effectively integrate over the full period in adding the results.

So far nothing too remarkable, although it does point up one way to perform decimation of a large FFT.

Now suppose each carrier in each mini-symbol is formed as the sum of four signal sets which are added or subtracted as shown in Figure 5. The complex carrier amplitudes of the carriers in the first mini-symbol are given by A_i and so on. These complex carrier amplitudes represent the coding of the carriers by QPSK or some higher order QAM

One signal component (J_1) always has the same sign in each symbol and is therefore conveyed just as we have explained above. Typically this would be coded by QPSK. The others are each added twice and subtracted twice. The receiver just outlined will therefore integrate them to zero and 'see' only the first component.

The other components can be retrieved by adding and subtracting appropriate combinations of the results of each of the small FFts we perform on each mini-symbol. Because of the linearity of the DFT we can add and subtract either the waveform samples or carrier complex amplitudes to taste. Clearly it is more efficient to add/subtract before the FFT at the modulator and after it at the receiver. The effective receiver operation is shown in Figure 6 where an equivalent signed 'integration window' is indicated for each signal-component set.

This kind of operation can be very neatly presented as matrix algebra and this is shown in Appendix A. The 'waveforms' in Figure 6 are, of course, Walsh functions.

The maxi-symbol has been shown divided into four mini-symbols. Any power of two could be used with the corresponding set of Walsh functions being chosen. It is probably also possible to use in fact any even number using an incomplete set of Walsh functions. In fact an odd number could be used if appropriate functions were selected.

So far little has been explained about the guard interval, shown in the diagrams as the shaded period. Suppose it takes the usual form of a repetition of the last fraction of the maxi-symbol; if we keep a guard duration of about one-quarter it would be a repeat of the last mini-symbol. There is however no need for it to be an integer number of mini-symbols.

When a received signal is delayed then that has the same effect as moving the receiver's integration window early with respect to the signal. The window then starts to overlap the guard interval. The information seen there replaces that lost at the tail end of the

window. Since harmonically-related sine waves remain orthogonal for any timing of the start of the integration window, provided its length equals the fundamental period, the signal is therefore received equally as well when delayed provided the delay does not exceed the guard interval. This is the well-known way in which the guard interval provides immunity to multipath in an OFDM system.

It is easy to see that reception of the J_i component will similarly benefit from this property. On the other hand, the other three will not and their reception will be progressively degraded for any delay at all. (A remedy for this will be suggested later). Clearly J_i is used as the rugged signal intended to convey the most important data required for both portable and rooftop reception.

EXTRACTING A PHASE AND AMPLITUDE REFERENCE

We have shown how sets of complex numbers J_i etc can be conveyed from transmitter to receiver, but not how data is mapped onto them. Clearly, each could carry data mapped onto a complex constellation. The difficulty is to extract phase and amplitude references for demodulation at the receiver.

The first clue is to note what is done in DAB. Here differential QPSK is used. The phase of each carrier in a given symbol period is the phase of that in the previous symbol plus a change of 0°. 90°, 180° or 270° depending on the 2 bits of data that the symbol has to carry. At the receiver, for each carrier, the difference is formed between the arguments of the complex amplitudes of the present and previous symbol. The received 2 bits are derived from the result. No explicit phase reference need be formed and the arbitrary phase shift introduced causes no difficulty (there is however a limit to the rate at which it may change). The price of this simplicity is a small penalty in thermal noise performance.

We can apply just the same method to map rugged channel data onto J_i so that differential demodulation occurs between maxi-symbols. Since these are of the same

duration as the regular symbols of DAB no problems are expected.

The higher bit-rate data intended for rooftop reception would be mapped onto the other complex-number sets using say 16-, 64- or 256-QAM. We propose the following trick to use the J_i as reference.

Consider one carrier in isolation, number 1 say.

At the transmitter we set $J_i = P_j e^{j\phi_1}$, where P_j is the (fixed) amplitude and ϕ_1 is the phase for this symbol, determined according to that of the previous symbol and the data to be conveyed in this 'rugged' channel.

At first let us ignore noise and suppose that the channel has complex frequency response H_1. At the receiver the result of the 'integration' for J will give $H_1 J_1 = H_1 P_j e^{j\phi_1}$.

Let the data to be conveyed by the K channel be mapped onto a normalised high-order QAM constellation whose value in this symbol has the complex value Q_{k1}. Suppose we set $K_1 = J_1 P_k Q_{k1}$ at the transmitter. P_k is a constant which lets us set the amplitude w.r.t that of the 'rugged' channel. At the receiver the result of the 'signed integration' for K is:

$$H_1 K_1 = H_1 J_1 P_k Q_{k1}.$$

If we divide this by the result of the J 'integrator' we get

$$\frac{H_1 K_1}{H_1 J_1} = P_k Q_{k1}$$

from which the transmitted data can be deduced without ambiguity. The effect of the channel's frequency response and any phase ambiguity have been removed.

Obviously the channel must remain static over the duration of one maxi-symbol for this to work. Indeed for the decoding of the 'rugged' channel which is conveyed as DQPSK the channel must be static over two maxi-symbols. Equally clearly the low-frequency component of local-oscillator phase noise is also cancelled by this technique.

What about thermal noise? The performance lies

between that of coherent and differential demodulation. If the amplitude of the K signal is set to be small compared with the J signal then the disturbance of the phase-and-amplitude-cancelling factor H_1J_1 by noise is proportionately smaller than the direct disturbance suffered by the received constellation.

What has been outlined can be extended to all the carriers and to the other 'sum-and-difference' channels L and M. Interestingly, although here applied to OFDM it could clearly be used with SC systems too.

The processing required is very simple -mostly add and subtract after doing four small FFTs, compared with the larger FFT necessary with conventional full-size symbols. The complex division would be needed in some form however the channel was measured.

A coder for deriving the doubly orthogonal signal as described above is shown in Figure 9. This comprises a digital data source 2 which could, for example, be an analogue to digital converter receiving an analogue video or audio signal and converting it to a digital signal. This source is shown, for the purposes of this example, with four outputs, 4, 6, 8, 10. The most important data which is required by all receivers, high and low quality, is provided at output 4 while less important data labelled other 1, other 2 and other 3 and only required by high quality receivers is provided on outputs 6, 8 and 10.

The most important data goes through a low order QAM mapper 12, in this case a differential QPSK mapper. This encodes pairs of the bits of the most important data by QPSK onto a constellation represented as a complex component by J_i. Data from outputs 6, 8 and 10 are of less importance and are sent to higher order QAM mappers 14, 16, and 18 which will map them onto, for example, a 16 QAM or a 64 QAM constellation in each case. The complex components onto which the various bits are mapped are represented as K, L, and M. The number of bits which can be represented by each of these components is determined by the order of QAM mapping being used.

In order that the outputs of each QAM mapper use the component J as a phase and amplitude reference. Each one is multiplied by J_i in multipliers 20, 22, and 24, respectively to yield K_i, L_i, and M_i. We thus have the four complex components $J_iK_iL_iM_i$ which form inputs to a Walsh/Hadamard matrix coder 26. This combines the components J, K, L, and M in the manner illustrated in Figure 5 to produce complex carrier amplitudes A_i, B_i, C_i, and D_i. Where i= 1 these will be the first carriers in each of the mini-symbols of Figures 4 and 5. The process will be repeated with additional bits of data until the required number of carriers for each mini-symbol have been generated.

At each stage the carriers A, B, C, and D output by the matrix coder are fed to a parallel to serial converter 28. Delays within this parallel to serial converter buffer the received inputs until the complete set of A, B, C, and D carriers making up the four mini-symbols have been received. These are then output to an inverse Fast Fourier Transform converter 30 which converts the carriers to a time domain signal which is sent to an appropriate modulator in preparation for transmission.

The Fast Fourier Transform coder is used as a relatively cheap way of implementing a Discrete Fourier Transform. The latter is what is required to perform the conversion and the FFT is merely one way of performing a DFT.

A corresponding receiver will have a decoder of the type illustrated in Figure 10. The time signal is the input to the decoder and corresponds to the time signal output by the inverse FFT 30 in figure 9. This time signal is input to a Fast Fourier transform unit 32 which reconverts to the frequency domain and outputs the converted data in serial form. This unit may perform a discrete Fourier Transform on the data. The data is the input to a serial to parallel converter 34 which, by a series of delays, is able to outputs on 4 output signals corresponding to A_i, B_i, C_i, and D_i of the type originally generated by the Walsh/Hadamard matrix coder 26 in Figure

9. As the FFT 32 will not start to output the D_i carriers of the fourth mini-symbol until three quarters of its total output time have elapsed the serial to parallel converter is correspondingly organised not to output data until this time. After this all the data has to be output in the following quarter of the total symbol time period.

The outputs form the inputs to an inverse Walsh/Hadamard matrix coder which regenerates the complex carriers J_i, K_i, L_i, and M_i which are, of course, the carriers originally modulated in the encoder with the digital data. The phase and amplitude reference is set by J_i. To recover the exact complex carriers coded at the encoder it is therefore necessary to divide the carriers K_i, L_i, and M_i by J_i in dividers 38, 40, 42. The J_i carrier is then passed to a DQPSK demodulator 44 whilst the K_i', L_i', M_i' carriers go to QAM demodulators 44, 46, 48. Demodulator 42 outputs the most important bits of data whilst the other demodulators output the other less important bits of data. If the data is used for an audio or video broadcast it can then be passed to an appropriate digital to analogue converter.

The above decoder is of the type which would be used with a high quality "roof top" receiver. In a less high quality receiver only the important bits of data would be used to reconstruct a signal. In such a case it would not be necessary to include the inverse Walsh/Hadamard matrix coder since integration over the whole time period would yield the J_i data which could then be fed to the DQPSK demodulator.

The arrangement shown will provide all the data simultaneously within a quarter of a symbol period and it will therefore be necessary to provide some buffering of output data to ensure continuity of signal reception.

REDUCING THE TIMING-ACCURACY REQUIREMENTS FOR ROOFTOP RECEPTION

The snag with the proposal outlined so far is that, while the 'rugged' J channel has the usual immunity to multipath (up to the guard-interval duration), the

other channels will degrade if there is any multipath or indeed mis-timing of the integration window. This isn't a cliff edge behaviour but is nevertheless undesirable. So how do we give some multipath protection to the fixed receiver?

A very simple proposal is to have full-length guard intervals between the mini-symbols. In effect we are returning to a classical full-size FFT and grouping together successive symbols to extract the phase reference. Drawbacks are thus the larger FFT and the longer timescale over which the channel and local-oscillator phase must remain stable.

Another idea is rather more complicated to explain. In effect it puts small guard intervals between the mini-symbols, gates all the integration windows and synthesises a special guard-interval extension before the first mini-intervals as shown in Figure 7.

It is easy to see how for delays less than the 'short' guard interval *all* integrations deliver the correct result and hence a 'rooftop' receiver using *K, L* and *M* as well as the *J*, and also portables will work.

For delays longer than the 'short' guard interval but shorter than the long one only reception of *J* can work. This is achieved by the guard-interval extension being a complicated sum-and-difference concoction of the bits of the rest of the symbols as necessary in order to cancel out those parts of the symbols which get included at inappropriate points in the integration. (See Appendix B for an illustration). A consequence of this is that the power in the extension may be greater than in the rest of the symbol.

The increase in power may not be too great a problem if we remember that the transmitter output back-off (OBO) will have to be set to keep intermodulation products (IPs) during the mini-symbols (and their associated short guard intervals) down to acceptably low level for the high-order modulation system used by the 'rooftop' receiver. If the power rises during the guard-interval extension so too will the IPs, but this part of

the waveform can only be used for reception of the rugged QPSK *J* channel. So the power rise may not be critical. It should be less so for a system with only 2 mini-symbols per maxi-symbol.

All of the methods discussed above are kinds of 'capacity-division multiplexing' where the capacity of an OFDM system (successive symbols, each able to carry 1 complex number value on each of say *n* carries) is shared out between various data channels carrying streams *J, K, L* ...

In 'pure' TDM, FDM or TDM/FDM, in a given symbol each carrier is carrying data from just one of these streams.

In the Doubly-Orthogonal proposal every carrier is treated similarly. Data from the *r* different streams are matrixed together to generate the complex carrier amplitudes of *r* successive mini-symbols. (The matrix in this case is a Hadamard matrix whose rows/columns are Walsh functions). One of the streams is coded using a rugged modulation system (DQPSK) and also serves as the phase and amplitude reference for the others. The matrixing takes place over groups of *r* mini-symbols, each group effectively constituting a maxi-symbol.

The Doubly-Orthogonal proposal is completely immune to the frequency-selective nature of the channel since each carrier is effectively 'sounded' by the rugged *J* channel conveyed on it. It does however require time-coherence of the channel and consistency of local-oscillator phase over the duration of the maxi-symbol.

Suppose it is desired to deal with signals having echoes delayed by up to 200μs.

If a guard interval of 200μs is inserted in front of each maxi-symbol, then just the *J* channel gets protected. The others have no protection against echoes, or, by the technique of guard-interval extension outlined above, they can be protected against short-duration echoes only. The total length of the maxi-symbol (for an overall guard-interval factor of 0.8) would be around 1 ms.

On the other hand, protection against echoes delayed by up to 200μs can be achieved for *all* data by

inserting a guard interval of 200μs in front of each *mini*-symbol. The price now is that keeping the same guard-interval factor of 0.8 makes the maxi-symbol length r ms, say 4 ms for 4 data channels.

MATRIXED FDM

The basis of the idea outlined below is the limitation of the channel impulse response to be essentially zero for t greater than the guard interval Δ. This is necessary anyway for the OFDM to work properly - as the delay increases beyond Δ, the orthogonality which isolates the carriers breaks down. While the effect of this is not immediately catastrophic in DAB, where rugged modulation is used, it appears to be more serious for higher-order modulation. Given that the channel impulse response is time-limited, it follows from a dual of the Nyquist sampling theorem that it can be adequately determined by a limited but sufficient number of samples in *frequency* of the channel frequency response. Thus if the guard interval, and the delay of the latest echo, is only a fraction of the symbol period we do not need to determine the frequency response of the channel at every OFDM carrier, but merely at a subset of them. The response at nearby carriers is thus to a degree *correlated*.

The proposed variant of capacity-division multiplexing uses FDM, where every rth carrier is ruggedly-coded and carries data channel J. The complex amplitudes of the in-between carriers are matrixed from the values of the higher-level-modulated channels K, L... The 'J carriers' are used as the phase and amplitude references for the others.

The basis of the proposal is thus matrixing in the frequency domain rather than the time domain.

Let the complex carrier amplitudes within a symbol be Y_i. If we have a division into r data channels then we arrange that every rth carrier carries a rugged data channel. For example take a simple case with r=3: We set

$$Y_{3i} = J_i$$

Now convey other, less important and less ruggedly coded date on the intervening carriers using the J_i as a phase and amplitude reference so we set:

$$Y_{3i-1} = J_i(K_i + L_i)$$

and

$$Y_{3i+1} = J_i(K_i - L_i)$$

Clearly this is a very simple example of matrixing.

To receive the data we do the appropriate additions and subtractions to undo the matrixing. That is easy. However, the result is made slightly more complex by the use of the J channel as the phase and amplitude reference. To appreciate this point we must take account of the channel response and the behaviour of the local oscillator in the receiver. (These could be lumped together but it is clearer to distinguish their effects if we treat them separately from the start).

Let the channel frequency response at the ith carrier be G_i (different for each carrier) while the instantaneous local-oscillator phase (w.r.t an arbitrary reference) is ϕ_N. This latter affects every carrier the same way. The received carrier amplitudes can thus be written as:

$$Y'_i = G_i e^{j\phi N} Y_i$$

The J data is received in the usual way by differential demodulation between symbols. Since a rugged form of coding is used this can tolerate some variation from symbol to symbol of the channel response and the LO phase. For the other data we form the following estimates:

$$\hat{K}_i = \frac{Y'_{3i-1} + Y'_{3i+1}}{2Y'_{3i}} = \left(\frac{G_{3i-1} + G_{3i+1}}{2G_{3i}}\right)K_i + \left(\frac{G_{3i-1} - G_{3i+1}}{2G_{3i}}\right)L_i$$

and

$$\hat{L}_i = \frac{Y'_{3i-1} - Y'_{3i+1}}{2Y'_{3i}} = \left(\frac{G_{3i-1} - G_{3i+1}}{2G_{3i}}\right)K_i + \left(\frac{G_{3i-1} + G_{3i+1}}{2G_{3i}}\right)L_i$$

It is easy to see that the phase noise term has been cancelled out (since we are using the same symbol as reference) while the channel response (including any temporal variation from symbol to symbol) is taken out provided that the response on adjacent carriers is the same. As the channel becomes more frequency-selective

the K and L channels will crosstalk and the phase and amplitude reference will be wrong. This could be a problem in some cases (e.g. with a strong echo delayed just less than Δ) even though the time-limitation of the impulse response has been honoured.

This system will work but is perhaps ill-proportioned for our application. It is very forgiving of temporal channel variation but at the price of placing too great a restriction on the variation of channel response with frequency. In particular the restriction of the channel impulse response to be essentially zero for t exceeding Δ is not a sufficient condition to ensure that it works.

Now suppose that we sound the channel regularly, once every frame of many symbols (as is done in DAB, for example). Let this produce results G'_i which are valid for the instant at which they were taken. Let us also take this measurement instant as the reference for the LO phase variation ϕ_N. Now write

$$G_i = \alpha_i G'_i$$

so that the α_i account for the time variation in the channel response at each carrier. At the receiver we first correct the received carriers using our measurement of the channel

$$Y''_i = \frac{Y'_i}{G'_i} = \frac{G_i e^{j\phi_N} Y_i}{G'_i} = \alpha_i e^{j\phi_N} Y_i$$

and then do the matrixing operations.

$$\bar{K}_i = \frac{Y''_{3i-1} + Y''_{3i+1}}{2Y''_{3i}} = \left(\frac{\alpha_{3i-1} + \alpha_{3i+1}}{2\alpha_{3i}}\right)K_i + \left(\frac{\alpha_{3i-1} - \alpha_{3i+1}}{2\alpha_{3i}}\right)L_i$$

$$\bar{L}_i = \frac{Y''_{3i-1} + Y''_{3i+1}}{2Y''_{3i}} = \left(\frac{\alpha_{3i-1} - \alpha_{3i+1}}{2\alpha_{3i}}\right)K_i + \left(\frac{\alpha_{3i-1} + \alpha_{3i+1}}{2\alpha_{3i}}\right)L_i$$

We note that the LO phase noise is still cancelled out. If the channel does not vary with time then the αs are all unity and cancel out, giving perfect reception whatever the frequency variation of the channel. If the channel varies with time then we require that variation to be a degree correlated between adjacent carriers over the time interval between measurements.

Clearly there is also a further noise penalty unless the channel-sounding results have been smoothed - which perhaps imposes stiffer requirements on temporal stability.

This system appears perhaps better-proportioned for our application. It is very forgiving of channel frequency-selectivity (indeed it is only necessary to restrict echo delays to less than Δ to satisfy the usual orthogonality constraints). The price is to place a restriction on the variation of channel response with time. LO phase-noise between symbols is still cancelled out.

A coder for use as a transmitter to code data in accordance with the matrixed FDM proposal is illustrated in Figure 11. As in the example described this is for a three carrier system. Clearly in real systems many more carriers will be used.

As in the coder of Figure 9 input digital data is separated into important data and other data. The important data is mapped onto a complex component by a DQPSK mapper 50 whilst the other data is mapped onto complex components K and L by QAM mappers 52 and 54.

Again the K and L components are normalised to the component J coded with the most important data for use as a phase and amplitude reference by multiplication in multipliers 56, 58. The products J(K+L) and J(K-L) are then generated in adder 60 and inverting adder 62.

These two products and the rugged component J then form the input to a parallel to serial converter. The connections to this are arranged as illustrated such that in serial form component J falls between the two other components. The whole is then input to inverse FFT coder 66 which outputs a time signal which can be prepared for transmission.

At a receiver for this time signal a decoder of the type shown in Figure 12 is used.

Firstly the time signal is reconverted to the frequency domain in an FFT unit 68.

The first thing that happens is that a reference signal extractor 70 extracts the reference

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signal carriers which are also sent with the time signal and derive the impulse response correction factor for each carrier. These correction factors are then fed to a one symbol delay 72 via a switch 74. Once all the correction factors are within the delay the switch is closed to form a loop between input and output of the delay so that the corrections will circulate within it. This correction factor is then used to divide the received time signal here designated as Y_i' to give the impulse corrected carrier Y_i''. These components are then input to a serial to parallel converter the outputs of which comprise the rugged component J and the two products J(K+L) and J(K-L). The components K and L are then extracted from these by adder and subtracter 78 and 80 and dividers 82 and 84. A DQPSK demodulator then extracts the bits of the most important data and two QAM demodulators extract the other data.

MORE COMPLEX EXTENSIONS

The reason why the simplest method above can fail in cases of interest (strong late echoes, as in SFN) is that $(G_{3i-1} + G_{3i+1})/2$ is a rather cheap-and-nasty interpolator for G_{3i}. If we could choose the matrixing of the data onto the in-between carriers in such a way that the corresponding term in the de-matrixing process was of the form

$$\frac{\text{(some rather better interpolator using } G_{3i+,-1} \text{ terms)}}{G_{3i}}$$

then the de-matrixing would work properly for all frequency selectivity subject to the restriction on the finite duration of the channel impulse response. There would be no need to add the complexity of the extra division of the method discussed above and as a consequence no restriction on the time variation of the channel between symbols.

It is possible to view the Doubly-orthogonal technique as a variant of this in as much as *if it is viewed over a maxi-symbol period* (instead of the mini-symbol over which the FFTs are actually performed) the spectrum contains isolated 'carriers' carrying *J* channel

information while every intervening 'carrier' carries a smeared-out matrix of the other information.

CAPACITY

The capacity of all the proposals based on time-, frequency-, time/frequency-division or doubly-orthogonal methods is essentially the same. In each case a certain proportion of the time/frequency slots is dedicated to a rugged channel for portables, with (D)QPSK modulation while the rest provide a higher bit rate with less ruggedness by using a higher-order modulation.

Suppose we use QPSK for *all* time/frequency slots, with a code rate of 0.62 and a guard-interval factor of 0.8. In This case the capacity (if we fill 7.6MHz of spectrum) is about 7.44 Mbit/s of 'rugged data'. At the other extreme if *all* time/frequency slots use 64-QAM with a code rate of 0.75 and guard-interval factor of 0.8 we have capacity of 27 Mbit/s (or 36 Mbit/s with 256-QAM). A practical proposal might lie between these extremes.

Consider the case when half the slots are devoted to each function. This could be a realistic example of the doubly-orthogonal case with synthesised guard-interval extension, or a spot example of TDM or FDM. The rugged channel would convey 7.44 ÷ 2 = 3.72 Mbit/s. The 'rooftop' channel would carry 13.5 Mbit/s (64-QAM) or 18 Mbit/s (256-QAM). These figures could be massaged by adjusting code rates or guard intervals.

The systems using TDM, FDM or TDM/FDM (but *not* the doubly-orthogonal or matrixed FDM method) can have the proportions trimmed fairly readily by simply adjusting the multiplex structure. Note also the possibility of increasing the guard-interval factor in the TDM case. In contrast the doubly-orthogonal method in effect always divides up the capacity in simple ratios.

APPENDIX A: MATRIX FORMULATION OF THE DOUBLY-ORTHOGONAL METHOD

A.1. General Notation

Each mini-symbol consists of say N complex samples of a time waveform which are related to N complex 'carrier' amplitudes by the Discrete Fourier Transform. In the following presentation I use lower-case e.g. x_i to represent complex waveform samples (time domain) and upper-case X_i for the complex 'carrier' amplitudes. The DFT can be written as follows:

DFT:

$$\begin{bmatrix} X_0 \\ X_1 \\ X_2 \\ \vdots \\ X_{N-1} \end{bmatrix} = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & 1 & 1 & \cdots & 1 \\ 1 & W^1 & W^2 & \cdots & W^{N-1} \\ 1 & W^2 & W^4 & \cdots & W^{2N-2} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ 1 & W^{N-1} & W^{2N-2} & \cdots & W^{(N-1)^2} \end{bmatrix} \begin{bmatrix} x_0 \\ x_1 \\ x_2 \\ \vdots \\ x_{N-1} \end{bmatrix},$$

where $W = e^{-j2\pi/N}$, and can be written more compactly as

$$\mathrm{X} = \mathrm{W}\mathrm{x}.$$

Inverse DFT:

$$\begin{bmatrix} x_0 \\ x_1 \\ x_2 \\ \vdots \\ x_{N-1} \end{bmatrix} = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & 1 & 1 & \cdots & 1 \\ 1 & W^{-1} & W^{-2} & \cdots & W^{-\overline{N-1}} \\ 1 & W^{-2} & W^{-4} & \cdots & W^{-\overline{2N-2}} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ 1 & W^{-\overline{N-1}} & W^{-\overline{2N-2}} & \cdots & W^{-(N-1)^2} \end{bmatrix} \begin{bmatrix} X_0 \\ X_1 \\ X_2 \\ \vdots \\ X_{N-1} \end{bmatrix}$$

or, in shorthand once more:

$$\mathrm{x} = \mathrm{W}^{-1}\mathrm{X}.$$

A.2. Basic Relationships for the Mini-Symbols

Each mini-symbol is described by a 'small' DFT e.g. $\mathrm{a} = \mathrm{W}^{-1}\mathrm{A}$. We can deal with the whole active maxi-symbol in one equation as follows:

$$\begin{bmatrix} \mathrm{a} \\ \mathrm{b} \\ \mathrm{c} \\ \mathrm{d} \end{bmatrix} = \begin{bmatrix} \mathrm{W}^{-1}\mathrm{A} \\ \mathrm{W}^{-1}\mathrm{B} \\ \mathrm{W}^{-1}\mathrm{C} \\ \mathrm{W}^{-1}\mathrm{D} \end{bmatrix} = \begin{bmatrix} \mathrm{W}^{-1} & 0 & 0 & 0 \\ 0 & \mathrm{W}^{-1} & 0 & 0 \\ 0 & 0 & \mathrm{W}^{-1} & 0 \\ 0 & 0 & 0 & \mathrm{W}^{-1} \end{bmatrix} \begin{bmatrix} \mathrm{A} \\ \mathrm{B} \\ \mathrm{C} \\ \mathrm{D} \end{bmatrix} = \hat{\mathrm{W}}^{-1} \begin{bmatrix} \mathrm{A} \\ \mathrm{B} \\ \mathrm{C} \\ \mathrm{D} \end{bmatrix}$$

and the corresponding inverse relationship:

$$\begin{bmatrix} \mathrm{A} \\ \mathrm{B} \\ \mathrm{C} \\ \mathrm{D} \end{bmatrix} = \begin{bmatrix} \mathrm{W}\mathrm{a} \\ \mathrm{W}\mathrm{b} \\ \mathrm{W}\mathrm{c} \\ \mathrm{W}\mathrm{d} \end{bmatrix} = \begin{bmatrix} \mathrm{W} & 0 & 0 & 0 \\ 0 & \mathrm{W} & 0 & 0 \\ 0 & 0 & \mathrm{W} & 0 \\ 0 & 0 & 0 & \mathrm{W} \end{bmatrix} \begin{bmatrix} \mathrm{a} \\ \mathrm{b} \\ \mathrm{c} \\ \mathrm{d} \end{bmatrix} = \hat{\mathrm{W}} \begin{bmatrix} \mathrm{a} \\ \mathrm{b} \\ \mathrm{c} \\ \mathrm{d} \end{bmatrix}.$$

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A.3. Applying the Walsh-Hadamard Mapping

At the transmitter the carrier amplitudes of the mini-symbols are derived from four other sets of complex amplitudes by a pattern of additions and subtractions that are in fact related to the Hadamard Transform:

$$\begin{bmatrix} A \\ B \\ C \\ D \end{bmatrix} = \begin{bmatrix} I & I & I & I \\ I & I & -I & -I \\ I & -I & -I & I \\ I & -I & I & -I \end{bmatrix} \begin{bmatrix} J \\ K \\ L \\ M \end{bmatrix}$$

where I is the $(N \times N)$ unit matrix. We can rewrite this as:

$$\begin{bmatrix} A \\ B \\ C \\ D \end{bmatrix} = H \begin{bmatrix} J \\ K \\ L \\ M \end{bmatrix}$$

so that the transmitted sequence for the maxi-symbol becomes:

$$\begin{bmatrix} a \\ b \\ c \\ d \end{bmatrix} = \hat{W}^{-1} \begin{bmatrix} A \\ B \\ C \\ D \end{bmatrix} = \hat{W}^{-1} H \begin{bmatrix} J \\ K \\ L \\ M \end{bmatrix}.$$

The inverse relationship applied at the receiver is (since $H^{-1} = H$):

$$\begin{bmatrix} J \\ K \\ L \\ M \end{bmatrix} = H^{-1} \begin{bmatrix} A \\ B \\ C \\ D \end{bmatrix} = H \hat{W} \begin{bmatrix} a \\ b \\ c \\ d \end{bmatrix}.$$

It is of interest to note the form of $H\hat{W}$:

$$H\hat{W} = \begin{bmatrix} W & W & W & W \\ W & W & -W & -W \\ W & -W & -W & W \\ W & -W & W & -W \end{bmatrix}.$$

A.4. The Relationship Between the Spectra of the Maxi- and Mini-Symbols

We could do a $(4N \times 4N)$ DFT on the whole retrieved sequence of $4N$ complex samples. In this case the relationship can be written:

$$Y = Vy$$

where y is the received sequence $\begin{bmatrix} a \\ b \\ c \\ d \end{bmatrix}$ and V is the $(4N \times 4N)$ DFT matrix:

$$\mathrm{V}=\frac{1}{\sqrt{4N}}\begin{bmatrix}1 & 1 & 1 & \cdots & 1\\ 1 & V^{1} & V^{2} & \cdots & V^{4N-1}\\ 1 & V^{2} & V^{4} & \cdots & V^{8N-2}\\ \vdots & \vdots & \vdots & \vdots & \vdots\\ 1 & V^{4N-1} & V^{4N-2} & \cdots & V^{(4N-1)^2}\end{bmatrix}$$

where the 'twiddle factor' V is now $e^{-j2\pi/4N}$.

Clearly $W=V^4$. It follows that every fourth row of V is (neglecting an overall scalar factor of 2) the fourfold repetition of a row of W. This in turn means that we can write an expression for every fourth carrier Y_0, Y_4, Y_8, ... as:

$$\begin{bmatrix}Y_0\\ Y_4\\ Y_8\\ \vdots\\ Y_{4(N-1)}\end{bmatrix}=\mathrm{Y}'=\frac{1}{2}[\mathrm{W}\quad \mathrm{W}\quad \mathrm{W}\quad \mathrm{W}]\begin{bmatrix}\mathrm{a}\\ \mathrm{b}\\ \mathrm{c}\\ \mathrm{d}\end{bmatrix}.$$

Compare this with

$$\begin{bmatrix}\mathrm{J}\\ \mathrm{K}\\ \mathrm{L}\\ \mathrm{M}\end{bmatrix}=\mathrm{H}\hat{\mathrm{W}}\begin{bmatrix}\mathrm{a}\\ \mathrm{b}\\ \mathrm{c}\\ \mathrm{d}\end{bmatrix}=\begin{bmatrix}\mathrm{W} & \mathrm{W} & \mathrm{W} & \mathrm{W}\\ \mathrm{W} & \mathrm{W} & -\mathrm{W} & -\mathrm{W}\\ \mathrm{W} & -\mathrm{W} & -\mathrm{W} & \mathrm{W}\\ \mathrm{W} & -\mathrm{W} & \mathrm{W} & -\mathrm{W}\end{bmatrix}\begin{bmatrix}\mathrm{a}\\ \mathrm{b}\\ \mathrm{c}\\ \mathrm{d}\end{bmatrix}.$$

By inspection $\mathrm{Y}'=\mathrm{J}/2$, i.e. the values of every fourth 'carrier' as measured over the long maxi-symbol period represent directly the 'rugged' J-channel information.

The values of the other in-between 'carriers' Y_i are less intuitively related to the K, L, and M data:

$$\mathrm{Y}=\begin{bmatrix}Y_0\\ Y_1\\ Y_2\\ \vdots\\ Y_{4N-1}\end{bmatrix}=\mathrm{V}\begin{bmatrix}\mathrm{a}\\ \mathrm{b}\\ \mathrm{c}\\ \mathrm{d}\end{bmatrix}=\mathrm{V}\hat{\mathrm{W}}^{-1}\mathrm{H}\begin{bmatrix}\mathrm{J}\\ \mathrm{K}\\ \mathrm{L}\\ \mathrm{M}\end{bmatrix}.$$

APPENDIX B: SYNTHESIS OF GUARD-INTERVAL EXTENSION

The idea, when using the 'doubly-orthogonal' method, of using

- short guard intervals (one per mini-symbol), to protect all data against short-delay echoes,

in combination with

- a synthesised guard-interval extension, to ensure that the ruggedly-coded *'J'-channel* data is always correctly received for delays less than the combination of the extension and the short guard interval,

was introduced without explanation above.

To understand this process we first revise the operation of conventional guard intervals in COFDM, as used for DAB for example. The DFT performed in the OFDM receiver in effect correlates the received signal against a set of sinusoidal (strictly, *complex exponential)* basis functions, performing an integrate-and-dump operation (in the form of a summation of samples) for each 'carrier' over a duration equal to that of the active symbol period T_a. The frequency spacing of the basis functions is $1/T_a$. It is a special property of the complex exponential basis functions that they remain orthogonal when any two are correlated for a duration T_a, whatever the time of the start of the T_a correlation (or 'integration') period. They also remain orthogonal if one of them is delayed, the correlation result (for each basis) function or 'carrier') simply having its argument changed - change of phase. This is only true provided that the signals being correlated are cyclic over the duration T_a. This is the reason for extending the symbols in OFDM by adding a guard-interval in such a way that a segment T_a long of the signal may be taken anywhere in a window of width equal to the total symbol length (T_a plus guard interval Δ) and throughout this segment every carrier remains continuous,

regardless of its modulation from symbol to symbol. In effect a section Δ long is taken from the *end* of the active symbol period and grafted in front at the beginning. In this way orthogonality is always preserved.

In the doubly-orthogonal proposal, correlations are performed over two (or more, depending on the number of Walsh functions used) successive active mini-symbol periods with a brief gap in between of duration equal to the short guard interval Δ_s. While the signal is delayed by no more than Δ_s each correlation over a mini-symbol independently delivers the correct result, just as for normal OFDM. The results from the mini-symbols can then be added and subtracted in turn to give the results for the *J*, *K*, etc. data channels provided by the 'doubly-orthogonal' method.

When the delay exceeds Δ_s, then the cyclic property of the signal is not maintained in either/any of the mini-symbols. However, by exploiting the linearity of (a) the DFT and (b) the process of deriving the J_i components by adding the DFT results, it is possible to synthesise a guard interval extension which contains appropriately-timed segments of the signal waveforms to cancel out (when the mini-symbol results are added) the inappropriate parts which now fall within the integration window. Of course we only have sufficient degrees of freedom to do this for the rugged 'sum' *J* data channel. The operation of such a synthesised guard-interval extension is illustration in Figure 8 where, for simplicity, a 2-channel example is shown with the mini-symbols containing the artificially-small number of 16 samples.

Incidentally, note the linearity means that a receiver for the *J* channel only could sum the received time-waveform data for the two mini-symbols (add across a mini-symbol shift-register delay) and then perform just one 'small' FFT per maxi symbol.

CLAIMS

1. A method for coding an orthogonal frequency division multiplex (OFDM) signal comprising the steps of dividing each OFDM symbol into a plurality of mini-symbols wherein carriers in each mini-symbol are formed from different portions of a plurality of signal sets, wherein the most important data is coded on a signal set which is substantially constant on corresponding carriers in the mini-symbol in an OFDM symbol and less important data is coded on signal sets that vary between corresponding carriers from mini-symbol to mini-symbol in dependence on predetermined functions wherein the signal sets used for less important data will sum substantially to zero across an OFDM symbol.

2. A method for coding an OFDM signal according to claim 7 wherein each mini-symbol includes a guard interval.

3. A method for coding an OFDM signal according to claim 7 or 8 wherein each symbol includes a guard interval extension.

4. A method for coding an OFDM signal according to claim 7 or 8 in which the most important data is coded into the first signal set using a first level of quadrature amplitude modulation (QAM).

5. A method for coding an OFDM signal according to claim 10 in which the less important data is coded into the other signal sets with a higher level of QAM.

6. A method for coding an OFDM signal according to any preceding claim in which the signal set coded with the most important data comprises a phase and amplitude reference for the other signal sets.

7. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal with data comprising

means for subdividing each OFDM symbol into a plurality of mini-symbols, means for coding corresponding carriers locations in each mini-symbol with different portions of the plurality of signal sets, wherein the most important data is coded on a signal set which is substantially constant for corresponding carriers in each mini-symbol and less important data is coded on signal sets that vary between corresponding carriers from mini-symbol to mini-symbol in dependence on predetermined functions.

8. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal according to claim 7 in which the predetermined functions comprises orthogonal functions.

9. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal according to claim 7 in which the coding means comprises an orthogonal function coding means.

10. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal according to claim 8 or 9 in which the orthogonal functions comprise Walsh functions.

11. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal according to any claims 7 to 10 comprising means for coding data onto the signal sets by Quadrature amplitude modulation.

12. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal according to any of claims 7 to 11 comprising parallel to serial conversion means for converting groups of coded mini-symbols to serial form.

13. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal according to any of claims 7 to 12 in which the signal set coded with the most important data comprises a phase and amplitude reference

for the other signal sets.

14. A method for coding an orthogonal frequency division multiplex (OFDM) signal comprising the steps of deriving a plurality of signal sets from the data to be coded wherein some of the signal sets are coded with the most important data and others are coded with less important data, assigning signal sets encoded with the most important data to every rth carrier forming an OFDM symbol, and assigning signal sets coded with less important data to intermediate carriers whereby each intermediate carrier is coded with data from a plurality of signal sets.

15. A method for coding an orthogonal frequency division multiplex (OFDM) signal according to claim 14 wherein the signal set coded with the most important data comprises a phase and amplitude reference for the other signal sets.

16. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal comprising means for coding a plurality of signal sets with data, wherein some of the signal sets are coded with the most important data and other signal sets are coded with less important data, means for coding each rth carrier forming an OFDM symbol with a signal set coded with the most important data, and for coding intermediate carriers with data from a plurality of signal sets including those carrying less important data.

17. Apparatus for coding an orthogonal frequency division multiplex (OFDM) signal according to claim 16 in which the signal sets coded with the most important data comprise phase and amplitude references for the other signal sets.

18. A method for coding an orthogonal frequency division multiplex (OFDM) signal comprising the steps of coding a first portion of the signal with the most

important bits of data and coding a second portion of the signal with bits of data of lesser importance wherein the first portion of the signal comprises a phase and amplitude reference for the second portion of the signal.

19. A method for receiving a signal coded in accordance with claim 18 comprising the steps of determining a phase reference for the second portion of the signal from the first portion of the signal.

20. Apparatus for decoding an orthogonal frequency division multiplex (OFDM) signal coded in accordance with the method of any of claims 1 to 6 comprising means for separating the signal into a plurality of mini symbols, each comprising a plurality of encoded carriers, and deriving from corresponding carriers in each mini-symbol a plurality of coded signal sets, at least some of the signal sets being encoded with the most important bits of data, and the other signal sets being encoded with data of lesser importance.

21. Apparatus for decoding an orthogonal frequency division multiplex signal coded in accordance with the method of claim 14 comprising means for decoding the most important bits of data from every rth carrier in received OFDM symbol and means for deriving signal sets coded with less important data from a plurality of the received carriers.

The Patent Office

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Databases searched:

UK Patent Office collections, including GB, EP, WO & US patent specifications, in:

UK Cl (Ed.N): H4M (ME, MFX, MX), H4P (PDX, PEE)

Int Cl (Ed.6): H04J (1/02, 3/00, 9/00, 11/00, 13/00), H04L (5/00, 5/06, 27/00)

Other: ONLINE: WPI, INSPEC

Documents considered to be relevant:

Category	Identity of document and relevant passage	Relevant to claims
A	GB2280571A (BBC) Whole Document.	1 & 7

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